IN THE UNITED STATES PATENT AND TRADEMARK OFFICE

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| In re Application of **Dent** | PATENT PENDING |
| Serial No.: **09/945,002** | Examiner: Mr. Kevin Burd |
| Filed: **31 August 2001** | Group Art Unit: 2611 |
| For: **Interference Cancellation in a CDMA Receiving System** | Confirmation No.:1823 |
| Docket No: **4015-980** | |

**DECLARATION OF PAUL W. DENT**

I, Paul W. Dent, hereby declare as follows:

1) On or before 12 October 2000, I conceived the invention tentatively entitled "Interference Cancellation in a CDMA Receiving System" while I was employed at Ericsson Inc. I prepared an Invention Disclosure describing my invention no later than 12 October 2000. A copy of the cover pages for that Invention Disclosure is attached as Exhibit 1.

2) Ericsson Inc. employs a standard review process regarding patenting an invention. The process generally includes filing of a Patent Disclosure by the Inventors, e.g., Exhibit 1, reviewing the disclosure by a patent review committee, and instructing a patent agent or attorney to prepare and file the application. On average, this process took approximately 6-9 months from initial conception to filing during the relevant time period.

3) My invention was approved for patenting by Ericsson Inc. in March 2001, and the Invention Disclosure was forwarded to an outside patent counsel working with Ericsson Inc. on or about 20 March 2001, with a request to prepare and file a patent application. A copy of the corresponding instruction cover letter to the patent counsel is attached as Exhibit 2 hereto.

4) A draft patent application covering the invention disclosed in the provisional and second Patent Disclosures was prepared by patent counsel and sent to by me for comments on or about 22 June 2001. A copy of the corresponding cover letter is attached hereto as Exhibit 3.

After revisions to the application based on my comments, the application was filed with the U.S.P.T.O. on or about 31 August 2001, receiving U.S. Application Serial No. 09/945,002.

5) The invention claimed in U.S. Application Serial No. 09/945,002, of which I am the named inventor, was conceived before December 2000 and pursued with reasonable diligence until the filing of the corresponding application on or about 31 August 2001.

I hereby declare that all statements made herein of my knowledge are true and that all statements made on information and belief are believed to be true, and further that these statements were made with the knowledge that willful false statements and the like so made are punishable by fine or imprisonment, or both, under §1001 of Title 18 of the United States Code and that such willful false statements may jeopardize the validity of the application or any patent issued thereon.

3 JULY 2006
Date

Paul W Dent
Paul W. Dent

EXHIBIT

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| For Legal Operation Use |
|---|
| Docket Number: EUSO |
| Date Opened: 2001- 0 |
| ☐DCP ☐ERA ☐ER ☐Other |

# Ericsson Inc. Invention Disclosure Cover Form

1. Invention Title: INTERFERENCE CANCELLING CDMA RECEIVING SYSTEM

2. Disclosure Submitted by (Add additional sheets if more than three inventors):

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3. Inventor No. 1 is part of a ☐Hosted ☐Non-Hosted organization (please check one).

4. Date invention conceived (mm/dd/yy): 10/12/2000 5. Date invention reduced to practice:

6. Identify (including dates) any past or anticipated disclosure outside the company, such as publication, offer for sale, actual sale, discussions with business partners, etc.: —

7. Invention made using government or non-Ericsson funding?: NO

8. Present or proposed use of the invention (identify products and dates): 3G or 4G SYSTEMS

9. Identify related invention disclosures of which you are aware: TV-20001012 "MOBILE PHONE DEVICES USING COHERENT MACHINERY"

10. Please attach to this cover form your invention disclosure, along with any other relevant documentation (see "IPR at RTP" Web site for additional information on writing disclosures).

The invention described in the attached invention disclosure is hereby submitted under my employment agreement with EUS

| Inventor's Full Signature | Date | Witnessed, read, understood and signed by | Date |
|---|---|---|---|
| (1) Paul W. Dent | 14 FEB 01 | (1) | |
| (2) | | (2) | |
| (3) | | (3) | |

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Patent Engineering

*Inventor: Submit to Kristen DeSimone, Dev. 1-2B, Cubicle #2421, x27711.*

Revision: 02/08/01

| ERICSSON | written by: | document no. | date: | signed:--------- |
|---|---|---|---|---|
| | Paul W. Dent | T/V20001012A | date:2000-10-12 | Paul W. Dent |

Inventor: Paul W. Dent

NOTE: THIS DISCLOSURE INCLUDES ALL THE MATERIAL FROM MY DISCLOSURE ENTITLED "MOBILE PHONE DEVICES USING COHERENT MACRODIVERSITY" BECAUSE THE THEORY IS CLOSELY RELATED. NEW TEXT DESCRIBING THE NEW ASPECT IS ADDED AT THE END. IT WOULD BE SENSIBLE FOR THE SAME ATTORNEY TO HANDLE BOTH.

# INTERFERENCE CANCELLING CDMA RECEIVING SYSTEM

## 1. FIELD

The invention describes a mobile communications network for receiving signals from multiple mobile susbcriber terminals and processing the signals to separate the mobile phone signals while eliminating cochannel interference therebetween.

EXEMPLARY CLAIMS: These exemplify the simplest impelementations of the concepts expounded herein, and the broadest claims

1. A mobile communications system for improving the performance of a communications service provided by a network of wireless base stations to a plurality of wireless subscribers, comprising:-

   -A receiver for receiving signals at a base station antenna site and digitizing the signals for numerical processing;

   -numerical processing for processing said digitized signals by correlation with time-shifts of a set of CDMA spreading codes and RAKE combining means to produce despread signal samples corresponding to information symbols with Intersymbol Interference;

   -channel estimation means for producing independent estimates of the complex amount of Intersymbol Interference in said despread signal samples between a signal sample corresponding to a symbol transmitted by any one of said subscriber terminals and a symbol transmitted by any other of said subscriber terminals, said estimates being assembled in a matrix of polynomials;

   -a symbol-rate processor for multiplying a vector of signal samples, each signal sample corresponding to one symbol transmitted by each of said subscriber terminals, by an Interference Suppression matrix formed using the adjoint of said matrix of polynomials in order to suppress interference from one subscriber terminal signal to another to produce interference-suppressed signal sample streams each associated with a subscriber terminal signal;

   -for each of said mobile subscriber terminal-associated signal sample

streams, an equalizer for processing said interference suppressed sample streams to decode information symbols from the associated mobile terminal while compensating for intersymbol interference between successive symbols transmitted from the same terminal.

2. A wireless base station for decoding symbols transmitted using Code Division Multiple Access from a number of wireless terminals, comprising:-

   -for each of said mobile terminals, a RAKE receiver for despreading the corresponding CDMA signal to produce a despread signal sample per information symbol;

   -decorrelation matrix processing operating on said information-symbol related samples to reduce interference from the symbols transmitted by one of said terminals to the signal samples corresponding to another of said terminals;

   -Viterbi Maximum Likelihood Sequence Estimation processing for decoding said information symbols from one of said terminals while compensating for intersymbol interference between successive symbols from the same terminal.

3. A system for providing coherent macrodiversity reception for mobile phone transmissions using a network of base stations, comprising:-

   -at each participating base station, a RAKE receiver for despreading each mobile phone transmission in the service area of the base station to obtain a sample stream comprising one complex sample per information symbol corresponding to each mobile transmission;

   -a backbone network for transmission of said sample streams from said base stations to a receive signal processing center while preserving their relative phases and amplitudes;

   -a matrix combiner for combining said sample streams such as to cancel interference between different mobile phone transmissions thereby producing samples streams which each depend only on the information symbols of a single corresponding mobile transmission;

   -an equalizer for each interference-cancelled sampled stream to decode information symbols while compensating for intersymbol interference from adjacent symbols of the same transmission.

4. The system of claim 3 in which the number of participating base stations is one.

5. The system of claim 3 in which said equalizer is a Viterbi Maximum Likelihood Sequence Estimator.

6. The system of claim 3 in which said backbone network transmits numerically coded complex samples.

7. The system of claim 3 in which said matrix combining uses an adjoint matrix instead of an inverse matrix.

8. The system of claim 3 further including the use of noise-whitening filters between the output of said matrix combiner and the input of said equalizer.

## 2. SUMMARY

A plurality of wireless subscriber terminals located within the service area of a network of wireless base stations transmit spread-spectrum coded signals bearing information symbols to be decoded in the network, the signals also facilitating the network stations determining multipath propagation channel characteristics, such as by the inclusion of pilot symbols or CDMA pilot codes in the signals. According to the invention, a base station comprises first a conventional RAKE receiver that correlates the signals received on a given channel frequency with shifts of the spreading codes used by the terminal and combines the correlations using multipath channel estimates to obtain RAKE-combined signal samples at the rate of one sample per information symbol for each terminal. The base stations futher pool their respective RAKE-combined sample streams in a central processor in the case that more than one base station or antenna is employed jointly to receive the terminal signals.

The central processor models the mutual interference in the RAKE-combined outputs between different terminal signals on the same radio channel by combining correlation coefficients between the known spreading codes with the channel estimates to obtain a matrix of Z-polynomials. The RAKE-combined sample streams are assembled as a vector of sample streams and the vector of sample streams is multiplied by a decorrelation matrix formed from the matrix of Z-polynomials in order to reduce interference from one terminal's signal to another terminal's signal to obtain substantially mutual-interference-free sample streams each corresponding to one terminal's signal but bearing Intersymbol interference between adjacent symbols of the same terminal signal. Individual processing of each terminal-corresponding signal then decodes information symbols while compensating for the Intersymbol Interference to produce soft-decoded information symbols at the input of an error-correction decoder.

The above invention can be employed with a single base station having a single receiving antenna; with a single base station site having multiple independent antennas such as cross-polarized antennas, or with multiple base station receiving sites.

## 3. BACKGROUND

The prior art of diversity communications comprises using more than one alternate communications means either simultaneously or by adaptive selection to improve communications quality or reliability. The alternate communications means can employ either different frequencies, different time slots or periods, different antennas or different polarizations or any combination of the above, being the known respectively as frequency diversity, time diversity, space diversity or polarization diversity.

The more usual form of diversity is receive diversity, in which the alternate communications means comprises alternate reception means for receiving the same transmittter. When the alternate reception means comprises a plurality of antenna elements forming an array, the outputs of which are coherently combined, the technique is also known as receive beamforming. When the multiple antenna elements are widely separated so that a signal received at different elements may fade in an uncorrelated fashion, even to the extent of being received at either one or another element and not both at the same time, diversity combining the antenna output signals cannot be equated to forming a narrow reception beam, but nevertheless has a large advantage for mitigating the effects of fading, and is termed "space diversity reception". These forms of receive diversity are

exemplified in the following U.S. patents, which are hereby incorporated by reference herein.

6,081,566 Method and apparatus for interference rejection with different beams, polarizations, and phase references
6,006,075 Method and apparatus for transmitting communication signals using transmission space diversity and frequency diversity
5,991,282 Radio communication system with diversity reception on a time-slot by time-slot basis
5,963,874 Radio station arranged for space-diversity and polarization diversity reception
5,878,093 Interference rejection combining with frequency correction
5,867,791 Up link macro diversity method and apparatus in a digital mobile radio communication system
5,499,272 Diversity receiver for signals with multipath time dispersion
5,481,572 Method of and apparatus for reducing the complexity of a diversity combining and sequence estimation receiver

Receive diversity according to the above patents relies upon knowledge of the relative phase and amplitude of the same signal received at different antennas. This may be determined for example by correlating the antenna outputs with known symbol patterns embedded in the signal for that purpose, known as channel estimation. Diversity reception using channel estimation is described for example in the above '272 patent. How to use the channel estimates in a computationally efficient manner to decode symbols based on joint processing of all the antenna signals is the subject of the above '572 patent.

Transmit diversity is the less usual form of diversity, in which the transmitter uses multiple antennas to communicate to receivers each having a single receiving antenna. When the multiple transmitting elements are in relatively close proximity and form an array, the technique is known as transmit beamforming. Transmit beamforming is described in the following U.S patents which are hereby incorporated by reference herein.

6,088,593 Two-way paging system and apparatus
5,940,742 Two-way paging system and apparatus
5,909,460 Efficient apparatus for simultaneous modulation and digital beamforming for an antenna array
5,848,060 Cellular/satellite communications system with improved frequency re-use
5,812,947 Cellular/satellite communications systems with improved frequency re-use
5,631,898 Cellular/satellite communications system with improved frequency re-use
5,619,503 Cellular/satellite communications system with improved frequency re-use
5,594,941 A cellular/satellite communications system with generation of a plurality of sets of intersecting antenna beams
5,642,358 Multiple beamwidth phased array
5,594,941 A cellular/satellite communications system with generation of a plurality of sets of intersecting antenna beams

In the above transmit beamforming patents, beamforming relies on the assumption that the relative phases and amplitudes of the propagation path from each antenna element to a given receiver are known, predeterminable characteristics of the array. In the above '941 patent to current Applicant, it is also disclosed that mobile receivers may provide feedback information to the transmitter using a reverse communications channel to assist it in the predetermination of these array characteristics. This technique is called

"Mobile-Assisted Beamforming".

When however the propagation paths from each antenna to each receiver comprise multiple propagation paths of different delay, and each has an amplitude and phase that varies with time, it is not possible to use the above methods of coherent beamforming. This usually arises when the multiple transmit antennas are widely separated. In that case, the prior art describes the non-coherent method of transmit diversity variously called SIMULCAST, MULTICAST, or TRANSMIT MACRODIVERSITY. These terms generally apply to using more than one transmitting base station on different sites to transmit the same information to a receiver. For example, the civil ground-to-air radio systems transmit from the ground to an aircraft using the three nearest ground stations. The three stations employ frequencies relatively offset by more than the highest audio frequency so that their beat notes are inaudible. Police radio systems also transmit using multiple base stations, but generally synchronise the frequencies so that the summation at the receiver of signals received from multiple transmitters simply looks like the multipath reception expected for UHF ground propagation in any case, but provides immunity against one or other transmitter being shadowed from the receiver by terrain obstruction. When analog speech modulation is used, care must be taken to ensure that the modulation is in time sync and not relatively delayed between the stations. On the other hand, if digital transmission is employed, it can be advantageous to arrange that different transmitters transmit the same information but with a deliberate delay of +/- one digital modulation symbol. An equalizer in the receiver is able to combine these relatively delayed signals which exhibit uncorrelated fading, thereby achieving diversity gain. This diversity gain is not obtained if the signals are not delayed by one or more symbol periods. Using an equalizer for diversity combining signals received from two or more transmitters is described in U.S. patent no 5,327,577 (Uddenfeldt et al, reissued 2 Feb 1999) which is hereby incorporated by reference herein.

Such non-coherent simulcast or macrodiversity transmission is also exemplified in the following U.S. patents, which are hereby incorporated by reference herein:-

6,104,933 Method and apparatus for control of base stations in macro diversity radio systems
5,930,248 Radio communication system selectively using multicast with variable offset time
5,883,888 Seamless soft handoff in a CDMA cellular communications system
5,845,199 Simulcasting system with diversity reception
5,724,666 Polarization diversity phased array cellular base station and associated methods
5,812,935 Cellular system employing base station transmit diversity according to transmission quality level

CDMA cellular systems most commonly employ macrodiversity, which is also known as soft handoff, in which a mobile station on the edge of two cells receives the same information transmitted from both cells during the transition period between receiving service from one cell and later receiving service from the other cell, as described in the above '577 and '888 patents. Non-coherent transmit macrodiversity for CDMA is also described in U.S. patent no. 5,940,445 to Kamin. Kamin transmits a signal and a one-symbol/chip delayed signal from two antennas. Kamin furthermore does not associated one transmit power amplifer with one antenna, but employs the matrix-PA technique first described by Welti in U.S. patent no. 3,917,998 . The matrix PA technique provides the advantage of pooling the power of all the transmitters so that one antenna may transmit more than the power of one transmitter if required, providing that the sum of the powers transmitted by all antennas does not exceed the sum of the available transmitter powers. The advantage of the matrix PA is however unrelated to the advantage of transmit macrodiversity, and in any case can only practically be

employed when all transmitters are on the same site. The current invention however may be employed even when the transmitters and antennas are not on the same site.

Non-coherent transmit macrodiversity is less effective in some important ways than coherent beamforming. Coherent beamforming enhances the received signal both by the summing of the transmitted powers of all cooperating transmit antenna elements and by the focussing of the transmitted beam towards the receiver, providing antenna gain in addition. Macrodiversity however provides only the former and not the latter. Coherent beamforming has the additional advantage of diminishing the interference transmitted in other directions. This allows the same frequency channel to be re-employed in different directions without interference, increasing communications capacity, which can be measured in units of voice channels per megahertz per square kilomter for example. In this method, the receivers for which the same channel is re-employed must be spaced by more than the beamwidth of the transmission beam. This can be many kilometers when the array of transmitting antenna elements is on the same tower.

With the current invention of coherent macrodiversity transmission however, using antennas on widely separated sites allows discrimination between cochannel users that are as close as a few meters or even centimeters to eachother, enabling for example three cellular phones in the same room to use the same cellular channel simultaneously.

Non-coherent macrodiversity by contrast increases the geographical spread of the interference from a given signal and reduces the reuse of the same channel. It nevertheless increases capacity over the capacity that would have been achieved without macrodiversity, as many receivers are located in the border regions between different transmitter service areas, when a uniform area distribution of receivers applies, and thus benefit from macrodiversity.

The research publications listed below have shown that wireless systems using multiple transmit antennas and multiple receive antennas can achieve high datarate capacity.


[1] E. Telatar, Capacity of Multi-antenna Gaussian Channels, AT&T-Bell Labs Internal Tech. Memo., June 1995.

[2] G. J. Foschini and M. J. Gans, On Limits of Wireless Communication in a Fading Environment When Using Multiple Antennas, Wireless Personal Communications, vol.6, (no.3), pp.311-35, Klu-wer Academic Publishers, March 1998.

[3] G. J. Foschini, R. A. Valenzuela, Initial Estimation of Communication Efficiency of Indoor Wire-less Channel, Wireless Networks 3, pp. 141-154, 1997.


The above references are hereby incorporated by reference herein. The background to the techniques described therein is similar to that of the current invention. An NxN matrix describing propagation from each of N transmit antennas to each of N receive antennas is essentially invertable to produce N separated channels on the same frequency that can each be used to convey a basic datarate, thereby achieving N times the datarate in total. However, in this art, the aim is to increase the datarate to a single reeceiver having the N receive antennas, and is achieved through the ability to jointly process the output of the N antennas at a single location. Typically, a receive matrix multiplication is performed to separate the N channels, which are then decoded using N, single-channel modems. The receive matrix multiplication can however only be performed if the N receive antenna signals are available at the single location.
In the current invention however, the N receive antennas are not colocated but may be miles apart, and each is associated with a different receiver.

The current invention therefore aims to achieve similar advantage without the increase of complexity entailed by equipping each with N antennas and receiver chains.

Capacity is increased beyond that achieved with non-coherent transmit macrodiversity if the geographical spread of interference from a given signal can be diminished. The current application discloses coherent macrodiversity, which can be employed when the transmitters have knowledge of the multipath propagation channel coefficients describing multiple, relatively delayed propagation paths to each of a number of receivers that receive different information using the same channel from the same transmitters, herein abbreviated to Channel State Information or CSI. The invention to be described provides the advantages of coherent beamforming, normally obtainable only with cosited antenna arrays, as well as the advantages of space diversity for overcoming terrain obstruction or uncorrelated fading, normally obtained only with widely spaced or non-cosited antenna elements. In the inventive coherent macrodiversity, the geographical spread of interference is reduced because the signals from multiple transmitters enhance one another by a maximum amount only at the intended receiver antenna location and do not enhance one another to the same degree in other locations even only centimeters away.

## 3. DESCRIPTION

The propagation from a transmit antenna to a receiver may be described by the following equation:-

$$R(i) = Co.S(i) + C1.S(i-1) + C2.S(i-2) \;....\; + C_{L-1}S(i-L+1) \quad ...(1)$$

where R(i) is the complex value of the received signal sample (with carrier frequency removed) at time instant i;

S(i) is the symbol transmitted at time i and Cj is the complex number describing the phase and amplitude of the propagation path with j-symbol periods of delay (relative to the shortest path, Co).

The combination of a signal with itself delayed in steps of one symbol or sample period may also be described by the use of the Z-transform. A discrete-time signal S that is delayed by one sample is denoted by $z^{-1}.S$, where "z" is the time advance operator and its reciprocal is the time delay operator.

Thus the polynomial $a + b\,z^{-1} + c\,z^{-2}$ times S means

$$aS(i) + bS(i-1) + cS(i-2).$$

Thus equation (1) can be written $R = C(z^{-1}).S$

where C is now the polynomial, in $z^{-1}$, with coefficients $Co....C_{L-1}$, R stands for the whole received signal sample stream and S stands for the whole transmitted symbol stream.

Denote by $C_{jk}$ the polynomial describing multipath propagation from antenna k to receiver j, with $R_j$ denoting the signal received at receiver j and $T_k$ denoting the signal transmitted by transmit antenna k. Thus the entire picture of propagation from all antennas to all receivers is described by the matrix equation

$$R_j = [C_{jk}].T_k$$

where [C] is now a matrix, each of whose elements is a polynomial in $z^{-1}$.

If we desire each receiver j to receive only an intended symbol stream $S_j$ then the transmitted signals needed to achieve that are given by

$$T_k = [C]^{-1}.S_j$$

Thus the problem at hand concerns how to invert a matrix [C] of Z-polynomials, and whether such an entity exists and is numerically well- or ill-conditioned.

Matrix theory describes the procedure for computing the inverse of a matrix as follows: First, transpose the matrix then replace every element by its cofactor to obtain a matrix known as the adjoint matrix. Then divide each element by the determinant of the original matrix to obtain the inverse matrix.

The determinant of a matrix is given by sums of products of its elements and is computable in a predetermined, systematic fashion. For example, for the 3 x 3 matrix

$$\begin{matrix} C11 & C12 & C13 \\ C21 & C22 & C23 \\ C31 & C32 & C33 \end{matrix}$$

the determinant is

$$C11(C22.C33-C32.C23)-C12(C21.C33-C31.C23)+C13(C21.C32-C31.C22)$$

The cofactor of element C11 is (C22.C33-C32.23), which is therefore the first element of the adjoint matrix. The first element of the inverse matrix is thus

$$\frac{(C22.C33-C32.c23)}{C11(C22.C33-C32.C23)-C12(C21.C33-C31.C23)+C13(C21.C32-C31.C22)}$$

When the C's are z-polynomials, it may be deduced that the numerator is a z-polynomial of twice the order of the C's while the denominator is a polynomial of three times the order. The inverse matrix comprises elements that have both a numerator z-polynomial and a denominator therefore. The signals to be transmitted must be processed by multiplication with respective inverse matrix elements and summed to produce the transmitted signals.

There is no problem in processing a signal by multiplication with a numerator polynomial, which just involves delaying the signal by multiples of a symbol period, weighting and adding, i.e. an FIR filter. The signal might conveniently be assembled into blocks, such as TDMA bursts, and a whole burst processed at once within digital signal processing memory before being output to the transmit modulator circuits. The processed burst is longer than the original by the length of the impulse response of the processing z-function, but can be prevented from overlapping adjacent bursts by allowing a guard time between bursts, as is common in TDMA systems such as the GSM cellular system.
In the case of a continuous signal such as CDMA for example, the signal can be segmented into blocks, processed in the same way as described above for a TDMA burst, and then the processed segments, which are now longer and overlap, are linearly superposed so that the extended tails add to the neighboring segments. It is acceptable for the impulse response tails to overlap neighboring segments so long as the intended receiver positions are the same for the neighboring segments. When the intended receiver positions are

different for neighboring segments, the tails of a previous segment overlapping a new segment would cause interference to be transmitted in the direction of the new receivers. In this case, adequate guard times should be used, or alternatively a means to allocate receivers to timeslots based on position can be used so that the spatial pattern of receivers served in adjacent timeslots is similar.

Processing a signal by a denominator z-polynomial is more problematic however. Such a z-function is an Infinite Impulse Response (IIR) filter. The impulse response tails off exponentially as a function of the successive powers of the poles of the z-function, i.e. the roots of the denominator polynomial. Thus if a root has a magnitude less than 1, the impulse response decays; if the root has a magnitude equal to one, the impulse response rings forever; and if the root has a magnitude greater than one, the impulse response grows. The latter however may be dealt with by processing the hypothesized signal block backwards in time to apply those roots that are greater than unity. Thus, the signal block is passed backwards through an IIR filter comprised of the reciprocal roots instead, which are less than unity and therefore give a decaying time-reversed impulse response. The roots that were already less than unity form an IIR filter which is applied in the forward time direction. It remains to determine what to do about roots that lie close to unity. Due to their slow decay, excessive latency in processing the signal exactly would be entailed. If the signal is not processed accurately, by truncating the IIR response for example, there is a risk of causing excessive interference overlap with adjacent blocks because the exact inverse C-matrix would not have been applied.

If the exact inverse C-matrix is applied to the information signals, the receivers will not only receive ONLY their intended information, but will receive it with multipath propagation cancelled already at the transmitter, since the resulting channel from the transmitter to the receiver is $C.C^{-1}$, which is unity.

Omitting division by the troublesome denominator polynomial is equivalent to multiplying each signal by the denominator polynomial, so the resultant channel from transmitting network to each mobile would be equal to that polynomial, which in the above 3x3 example was three times the length of the individual multipath channel polynomials $C_{jk}$. The receivers can have an equalizer to decode symbols in the presence of normal amounts of multipath propagation delay, but perhaps not three times that amount. Therefore omitting the entire denominator may produce a net channel that exceeds the receiving equalizer's capability. However, one third of the factors in the denominator can be omitted and this is equivalent to multiplying by a polynomial of 1/3rd the order of the denominator, thereby creating an artificial multipath channel only of the same length as the original multipath channel. According to one aspect of this invention therefore, the 1/3rd of the denominator factors which are omitted are chosen to be those corresponding to roots of magnitude closest to unity, which are the terms that cause the most slowly decaying impulse responses as well as peaks in the frequency spectrum of transmitted energy. Preferably, the roots having logmagnitude closest to zero may be selected as the roots of magnitude closest to unity.

In summary of the above improved system of communication therefore, the following is proposed:

A transmit processor is provided with NxNxL channel coefficients describing the NxN multipath channels of impulse response length L to each of N receivers from each of an equal number N of antennas available for transmission.

Treating each set of channel coefficients as a z-polynomial of order L and as an element of an NxN square matrix of such polynomials, the transmit processor

forms the NxN adjoint matrix whose elements are polynomials of order (N-1)L, and a determinant polynomial of order NL.

The determinant polynomial is factorized to determine the NL roots and the L roots having smallest absolute value of logmagnitude are discarded, leaving (N-1)L roots forming a reduced determinant polynomial of order (N-1)L . The signals for transmission are then processed through FIR filters of order (N-1)L corresponding to the adjoint matrix polynomials and through an IIR filter of order (N-1)L corresponding to the reduced denominator polynomial as follows:

Information blocks intended to be transmitted to different receivers are assembled in the transmit processor. Each information block is passed in the reverse time direction through an IIR filter formed from the reciprocals of roots of the reduced determinant that have a magnitude greater than unity and also in the forward time direction through an IIR filter formed from the reduced determinant roots of magnitude less than unity to obtain IIR-processed blocks.

The IIR processed blocks are then FIR processed by matrix multiplication with the adjoint matrix polynomials to obtain transmit blocks. The transmit signal blocks may be further filtered to restrict transmission bandwidth and upsampled and converted to continuous time signals. The continuous time signals are then I/Q modulated on a radio frequency carrier, amplified in respective transmit power amplifiers and transmitted using respective antennas. Since IIR, FIR and bandwidth-restriction filtering are all linear processes, they may be carried out in any order.

The above-mentioned polynomial lengths apply to the case where the three transmitting antennas are on the same sites or nearby sites.
The modulators, transmit amplifiers and antennas need not be on the same site but can receive their transmit blocks from a common transmit processor, by optical fiber landline for example. Different sites must however synchronise the frequencies of their radio frequency carrier generators to the same stable reference, using GPS for example. The relative phases must be at least as stable as the changes in multipath propagation channel coefficients, which means that their relative frequency accuracy should be better than the Doppler frequency caused by relative motion between the transmitters and the receivers. This ensures that the mechanism used to keep the transmit processor provided with up-to-date and accurate channel coefficients is also sufficient to compensate for any phase difference or phase drift between sites.
When the three transmitting antennas are on different sites far apart, the polynomial orders may be somewhat higher, as will be further discussed below.

A system according to the above description is shown in figure 1. Information symbol streams intended respectively for receivers RX1,RX2 and RX3 are input to numerical baseband processor (4). The processor (4) generates complex number streams having a real or in-phase part I and an imaginary or quadrature part (Q) representative of the information symbols modulated on to a carrier. Using knowledge of the multipath propagation polynomials Cij(z), the baseband processor then forms weighted combinations of the complex number streams and themselves delayed. The complex combination sample streams are then converted to continuous time waveforms using DtoA convertors and upconverted to a desired radio frequency channel using respective quadrature (I/Q) modulators (3A,3B,3C). The upconverted signals are then amplified to desired transmit power levels using respective Power Amplifiers (2A,2B,2C) and transmitted by respective antennas (1A,1B,1C). The transmitted signals propagate to receivers RX1,RX2,RX3 by way of the multipath channels Cij. Receivers RX1,RX2,RX3 thereby receive radio frequency signals modulated only with respectively intended information symbols as originally input

to baseband processor (4). In figure 1, antennas 1A,1B,1C may be on the same mast or tower on the same site or even on different towers at different sites. In the latter case, the connection from the outputs of baseband processor (4) to the inputs of modulators (3A,3B,3C) may be made by auxiliary intra-network communications means, such as optical fibre links, wireline or microwave links, always observing the necessary precautions to maintain the phase coherency between the signals across antennas. Various means to accomplish this coherency requirement are disclosed in the above-incorporated '941 patent to applicant.

Figure 2 shows more detail of baseband processor (4).
Signals for transmission to respective receivers are first IIR filtered by the same IIR filter, in this exemplary implementation, the IIR filter being the determinant z-polynomial of the propagation matrix reduced in order by omission of factors corresponding to roots close to the unit circle. The number of roots omitted may be anything from none up to the receiver's equalizer capability. Each omitted root requires the receiver to deal with one extra symbol period's worth of multipath time dispersion. If different receivers have different equalizer capabilities, the IIR filters 10A,10B and 10C could be correspondingly different, with different numbers of factors omitted. The IIR filters may also differ by factors intended to optimize communication efficiency separately for each receiver, as will be further discussed below.

After IIR filtering, the array of FIR filters (20) processes the signals. Each signal is processed by a corresponding row of FIR filters (20). The output signals from FIR filters (20) are summed down columns, indicated by the + sign at the junction of the line from one output to another. The summed outputs are then the signals to be modulated and transmitted from respective antennas.

In the case where the receivers are the same and IIR filters 10A,10B...10C are the same, the net channels by which respective signals reach respective receivers are identical, despite different propagation coefficients. In effect, the transmit preprocessing compensates exactly for the different channels. It is well known in the trivial case of N=1 (one antenna transmitting to one receiver) that precompensation for the propagation channel at the transmitter is not optimum. Such precompensation would use more transmit power to transport frequency components lying in a bad frequency region of the channel, having high path attenuation, and less power to transport frequency components in a low attenuation region. An optimum prefiltering would do the reverse, transmitting most of the transmitter power where the propagation was good and not wasting power where the propagation was bad. The known optimum solution for the single transmitter, single receiver case is called the "waterfilling" solution.

When received signal power varies with time instead of with frequency, i.e. so-called flat fading, the prior art of CDMA systems such as IS95 teaches that rapid power control should be used to attempt to keep the received signal levels more or less constant. However, the waterfilling solution for determining how transmitted energy should be distributed over the frequency spectrum may be applied to temporally varying channels showing that it is better to transmit less power when the channel is bad and more when the channel is good. Temporal waterfilling is the opposite of the power control technique propounded in prior art CDMA systems.

When received signal energy is a two dimensional function of both frequency and time, two-dimensional waterfilling should ideally be employed to determine the energy to be transmitted in each frequency subband at each instant.

The waterfilling solution has the characteristic that no power is transmitted in regions of the spectrum where the received signal is instantaneously below a threshold signal to noise ratio. Producing a spectral null over a broad spectral region however implies high order bandstop filtering, which may be impractical or produce time dispersion in excess of a receiver equalizer's

capability to compensate. The current invention avoids these disadvantages.

The principles for optimum prefiltering according to the invention for the multiple antenna and multiple receiver case will now be derived.

Instead of describing the propagation channels by z-polynomials, they could be described by frequency response functions Cjk(w). The frequency response function can be obtained from the z-polynomial simply be substituting exp(jwt) for z or exp(-jwt) for $z^{-1}$. The frequency response functions can be evaluated at each spot frequency to obtain a propagation matrix of complex numbers at the spot frequency. The matrix of complex numbers can be inverted by computing the adjoint matrix and the determinant to form the baseband processing required to transmit that spot frequency component of each signal to respective receivers in an optimum manner. This can be repeated for each spot frequency component to obtain the baseband processing required for each spot frequency component and thereby a set of inverse frequency response functions which form a matrix for pre-processing the transmit signals.

The spot frequency case has also been formulated in the above-incorporated '941 patent to Applicant for the case where the number of antennas (M) exceeds the number of receivers (N), i.e for the overdimensioned case.

If Cjk(w) is an N x M matrix of complex numbers describing the phase and amplitude of the path from antenna k to receiver j at a spot frequency w, and Sj(w) is the frequency component w of the information symbol stream Sj intended for receiver j, then the transmit signals Tk(w) that should be transmitted from antenna k in order to ensure that each receiver receives only its intended signal component, while minimizing the total transmit power needed, are given by:

$$Tk(w) = C^{\#}[C.C^{\#}]^{-1}Sj(w)$$

By evaluating $C^{\#}[C.C^{\#}]^{-1}$ at each spot frequency 'w' a matrix of transmit preprocessing frequency response functions is found. These can be converted to a set of impulse responses by Fourier Transforming them if desired, so that the preprocessing of the signals Sj can take place in the time domain with FIR filters. However, it is possible that, in attempting to invert $[C.C^{\#}]$ at certain spot frequencies w, it is found that its determinant is zero or near zero. This would signify one of two problem situations: Either that there is an unavoidable null in the frequency response of the channel from all antennas to at least one of the receivers, or that the propagation channels from each of the antennas to two or more receivers are identical, causing the matrix to be rank-deficient. As in the z-polynomial formulation, zeros in the determinant of a matrix may be avoided by omitting that factor, corresponding now to a zero root, from the determinant. More specifically, roots in the frequency domain that lie on or close to the the imaginary axis represent "high-Q" poles and may be removed by multiplying the frequency response functions $[C.C^{\#}]^{-1}C^{\#}$ with a zero term in the numerator having the same root. While this is in practice close to optimum, as in the time-domain formulation, this would result in all receivers experiencing the same net propagation channel, which seems unexpected and therefore likely to be non-optimum. Instead therefore, the following procedure is proposed:

When the spot-frequency inverse matrix has been computed (the adjoint matrix elements and the determinant can be kept separate to avoid attempting to divide by a zero determinant), the net spot-frequency power attenuation factor from the M-antenna diversity transmitter to each receiver may be computed by summing the squares of the magnitudes of the adjoint matrix elements down columns. The squared magnitude of the determinant is then divided by the column sum to

obtain a net power attenuation factor from the transmitting system to the receiver. Repeating that for all spot frequencies gives the power attenuation frequency response function from the transmitting system to each receiver. Now the N=1 theory can be applied to determine how a signal for a given receiver should be prefiltered for transmission through the asscociated channel having that determined net power/frequency response.

Prior art theory for N=1 gives the optimum transmit power spectral shaping P(w) for transmitting information through a channel with a given power spectral shape H(w) as "the waterfilling solution". The waterfilling solution considers pouring water onto an object of shape 1/H(w) (with walls at the band edges) until the total amount of "water" contained is equal to the total power available. The function P(w) then equals 1-Ho/H(w) where Ho is the reciprocal of the level of the water. No power is transmitted in regions of the band still not submerged, i.e. where H(w) < Ho. The solution thus depends on the total power available relative to the receiver noise and the mean path attenuation, i.e. on the mean SNR that can be created at the receiver. Given a desired datarate and a channel bandwidth, the mean SNR needed to sustain that datarate can be determined, and the water-pouring solution for that SNR used. The water-pouring solution results in zero power transmission in regions of the spectrum where the SNR is less than a threshold. It is complicated to realise a transmit prefilter with abrupt cut-off at the boundaries of such regions, and these prefilters moreover have long impulse responses, requiring complex equalizers at the receiver. A compromise filter that centers single zeros of the form $1+az^{-1}$ on the null-regions can be used instead, which is nearly equivalent to the above-proposed time-domain solution of deleting determinant polynomial roots close to the unit circle. The difference is that, in the just-described approach, the position "a" of the zeros may be adjusted slightly away from the true determinant zeros in order to flatten the net frequency response from the diversity transmitting system to each receiver separately, taking account therefore of the effect of each column of adjoint matrix elements, thereby satisying the intuitive expectation that the net channels should not turn out to be identical for each receiver.

The frequency-domain solution and the time-domain solution are in fact identical as the matrix

$$C(w)^{\#}[C(w).C(w)^{\#}]^{-1}$$

is just the matrix

$$C(z^{-1})^{\#}[C(z^{-1}).C(z^{-1})^{\#}]^{-1}$$

with $z^{-1}$ set equal to exp(-jwt), where the symbol # for matrices of z-polynomials means conjugate transpose and end-to-end (time-) reversal of the polynomials. An example of the correct way to form the time-reversed conjugate transpose of a matrix of z-polynomials is given in figure 11.

Therefore the procedure can simply comprise determining the matrix C of z-polynomials as before, and then forming the matrix of double-length (2L) z-polynomials:

$$C(z^{-1}).C(z^{-1})^{\#}$$

where # means time-reversed, conjugate transpose as defined in figure 13.

The above matrix is inverted as before by computing the adjoint matrix of z-polynomials of length 2L(N-1) and the determinant z-polynomial now of length 2LN, and then multiplying the adjoint matrix by $C(z^{-1})^{\#}$ to obtain a matrix of length (2N-1)L polynomials. The attenuation/frequency curve from the transmitter

to each receiver can now be determined by letting z = exp(jwt) and summing down matrix columns the squares of the moduli of the elements, and repeating for each spot frequency. The square of the determinant modulus is then divided by the row sums. For each receiver, the resulting attenuation/frequency curve may be examined to locate the highest L attenuation peaks. FIR zeros of the form $1 + az^{-1}$ are then placed on the attenuation peaks to either flatten the frequency response, which is close to optimum, or to over-flatten the response by turning the peaks into troughs, which is even closer to the optimum water-pouring solution. An approximate rule-of-thumb for over-flattening would be to choose a zero to turn a peak originally X decibels above the mean to a trough X decibels below the mean. The use of no more than L zeros ensures that the receivers do not need to equalize more than L symbol periods of net time dispersion. The result of applying the zeros to the signal will not however flatten the transmitter-receiver frequency response, as that is flat already when the zeros are not used; rather, it is the curve of transmit power versus frequency which is flattened across the spectrum, resulting in notches in the frequency response from the transmitting system to the receiver. These notches appear to the receiver to have been caused by multipath propagation with a channel equal to the product of the applied "flattening" zeros.

There is essentially no difference in the transmitter structure for the above solution as compared to figure 2, except for the choice of the FIR matrix polynomials and the IIR determinant polynomials, and prefilters 10A,10B,10C now also have an FIR element composed of the above-mentioned flattening zeros, if they do not exactly annihilate factors of the determinant. Note that if a determinant polynomial has a pole exactly on the unit circle, giving an infinite attenuation peak, then the pole would be annihilated by a flattening zero also exactly on the unit circle, i.e. $|a|=1$. Likewise poles close to the unit circle giving high attenuation peaks would be flattened by placing a zero nearby. If the zero was placed exactly on the pole, thereby annihilating it, the solution is identical to deleting the L poles of the determinant closest to the unit circle.

Figure 3 is a flowchart for a control processor located in the network for determining the FIR and IIR transmit preprocessor coefficients based on channel state information (CSI). At step 100, the CSI information is updated to reflect latest estimates of the donwlink channel z-polynomials Cij.

At step 101, a test is performed to determine whether the transmitter is trying to transmit to a greater, lesser or equal number of receivers N compared to the number of antennas M at its disposal. If N>M for this implementation the requirement cannot be satisfied and an error must have occurred. Error recovery procedures at step 102c can for example comprise selecting to serve only the first M receivers and not the remaining N-M. If N=M, the matrix C is square and it is effectively inverted by computing at step 102a the adjoint matrix of polynomials of order (N-1)L and the determinant polynomial of order NL. Then at step 103 the determinant polynomial is factorized, i.e. its roots are found and sorted in order of magnitude. At step 104, the L roots and factors which are closest in magnitude to unity are discarded. The number L of roots and factors discarded may be different if different receivers have different equalizer capabilities, which the transmitter would be informed about in advance.

If the number of receivers N is less than the number of transmitting antennas or sites M, then the extra degrees of freedom are used to optimize performance at step 102b, which computes the matrix $C.C^{\#}$ of polynomials of length 2L. Then the adjoint matrix of this matrix is computed as is its determinant. The adjoint matrix polynomials are of length 2L(N-1) and the determinant polynomial is of length 2NL. Other than that however, the procedure continues to the same step 103 as for N=M. After deleting the L roots closest to the unit circle, the determinant polynomial is now of length

(2N-1)L. Steps 105 and 106 are then performed with the resulting adjoint matrix and reduced determinant polynomial by the baseband processor of figure 2.

If it is desired to perform optimum transmit spectral shaping for each receiver separately, the flow chart of figure 4 may be used.

In figure 4, steps 103-105 of figure 3 have been repaced by steps 103a-105a of figure 4. At step 103a, the best achievable net attenuation versus frequency response from the transmitting system to each receiver is computed. It is then desired to warp the transmit spectrum for each receiver to direct more power to regions of the spectrum with more favorable attenuation and would have used less transmitter power and less to regions of the spectrum that suffer greater attenuation and would have used more transmitter power. The water-pouring algorithm is the optimum and could be used at step 104a for this if not considered too complicated. Otherwise the simpler approach of adding flattening or overflattening zeros can be used. In the case of high-Q determinant poles, i.e. roots very close to the unit circle, a flattening zero may be placed exactly over the pole to annihilate it. Instead of adding a zero then, a pole is annihilated from the determinant instead. On the other hand if one of the L poles closest to the unit circle is a low-Q pole, the attenuation frequency response may not show a peak exactly on the pole frequency but will be displaced due to the influence of the adjoint matrix FIR polynomials. In that case a zero is centered on the displaced peak and does not annihilate the nearby pole. Steps 105a and 106 are then performed by the baseband processor of figure 2.

An example will now be given for the case of three transmitting antennas on different sites far apart.Figure 5 shows three base stations that for simplicity have been sited at the corners of an equilateral triangle with 10km sides. The three sites are jointly supporting communications with multiple receivers, of which three are located at m1,m2 and m3 as shown, and it is desired to use the same channel frequency simultaneously if possible for all three in order to increase capacity according to this invention.
The base stations transmit an exemplary symbol rate of 13MHz/48, which is the symbol rate of the GSM/EDGE cellular system. The spatial spread of such a symbol is given by dividing the speed of light by the symbol rate, giving 1107 meters. The distances between the mobile receivers and the base stations expressed in terms of symbol wavelengths is shown in the table below:

| | TX1 | TX2 | TX3 |
|---|---|---|---|
| RX1 | 3.10636 | 7.1187 | 6.10841 |
| RX2 | 5.58839 | 5.53673 | 4.56929 |
| RX3 | 4.65383 | 4.66078 | 6.67428 |

If the channel for each of these nine paths were of equal amplitude and phase and were line of sight with no multipath, the z-matrix relating received signals R1,R2,R3 to transmitted signals T1,T2,T3 would thus be:

$$\begin{matrix} Z^{3.11}R1 \\ Z^{4.56}R2 \\ Z^{4.66}R3 \end{matrix} = \begin{pmatrix} 1 & Z^{-4} & Z^{-3} \\ Z^{-1} & Z^{-1} & 1 \\ 1 & 1 & Z^{-2} \end{pmatrix} \begin{matrix} T1 \\ T2 \\ T3 \end{matrix}$$

where the common fractional symbol delay parts of each row and a common integer power of Z have been transferred to the left into R, such that the powers of z remaining are simplified to the lowest integers, representing

whole symbols of delay.

Maintaining the simplifying assumption of equal phase and amplitude on all nine paths, the adjoint of this matrix is:

$$\begin{matrix} (Z^{-3}-1) & (Z^{-3}-Z^{-6}) & 0 \\ (1-Z^{-3}) & (Z^{-2}-Z^{-3}) & (Z^{-3}-Z^{-2}) \\ 0 & (Z^{-4}-1) & (Z^{-1}-Z^{-5}) \end{matrix}$$

and the determinant polynomial is $-1 + z^{-3} + z^{-4} - z^{-7} = -(1-Z^{-4})(1-Z^{-3})$

The determinant has all seven roots on the unit circle at

$z = 1$ (two roots)
$z = -1$
$z = j$
$z = -j$
$z = \exp(j120^0)$
$z = \exp(j240^0)$

Each root represents a frequency at which infinite attenuation can arise between the transmitting system and the receiving system, so it is inefficient to attempt to convey energy at those frequencies to the receivers. To avoid this, all seven roots on the unit circle should be annihilated by zeros in the numerator, which is the same as deleting the roots of the determinant. This could cause the receiver equalizer to have to deal with a length 7 channel, i.e. 7 symbols of time dispersion. All the adjoint matrix elements however share at least one root with the determinant that can be annihilated. Cancelling the factor $-1+Z^{-1}$ from both the adjoint matrix and the determinant polynomial leaves:

$$\begin{matrix} (1+z^{-1}+z^{-2}) & -Z^{-3}(1+z^{-1}+z^{-2}) & 0 \\ -(1+z^{-1}+z^{-2}) & -z^{-2} & z^{-2} \\ 0 & (1+z^{-1}+z^{-2}+z^{-3}) & -z^{-1}(1+z^{-1}+z^{-2}+z^{-3}) \end{matrix}$$

and the determinant is now $(1-z^{-4})(1+z^{-1}+z^{-2})$

Omitting to divide by the 6th order reduced determinant means that the receivers will receive their signals modified by a 6th order FIR filter, and their equalizers must be able to deal with 7 symbol periods of time dispersion.

The above simplified example gave rise to many determinant roots on the unit circle due to the assumption of equal amplitude paths. If the paths are not equal, but exhibit a single propagation path of amplitude and phase given by the complex numbers Cij, then the propagation matrix becomes instead:

$$\begin{matrix} Z^{3.11}R1 = \\ \\ Z^{4.56}R2 = \\ \\ Z^{4.66}R3 = \end{matrix} \begin{pmatrix} C11 & C12.Z^{-4} & C13.Z^{-3} \\ & & \\ C21.Z^{-1} & C22.Z^{-1} & C23 \\ & & \\ C31 & C32 & C33.Z^{-2} \end{pmatrix} \begin{matrix} T1 \\ \\ T2 \\ \\ T3 \end{matrix}$$

The Cij may be complex numbers that are changing due to Rayleigh fading.

If the paths are in addition multiple paths, the Cij can be z-polynomials as in the earlier examples. The additional powers of $Z^{-1}$ that are now attached to the Cij account for the extra propagation delay due to the substantial path differences that exist in the case of widely separated sites.

As an example, the above Cij were each chosen to be second order polynomials with three random Gaussian complex coefficients, representing multipath channels of three, synbol-spaced rays. A typical frequency response plot of the determinant polynomila is shown in Figure 7 plots the determinant polynomial and the flattened polynomial obtained by deleting the four roots closest to the unit circle. The coefficients of the example polynomial are given in the table below:-

| | REAL | IMAG |
|---|---|---|
| A(1)= | -0.01492 | 0.01770 |
| A(2)= | 0.01484 | -0.02824 |
| A(3)= | -0.02419 | -0.08202 |
| A(4)= | 0.00808 | -0.01929 |
| A(5)= | -0.04147 | -0.15490 |
| A(6)= | 0.36312 | 0.11521 |
| A(7)= | 0.44006 | 0.19125 |
| A(8)= | 0.58783 | 0.88261 |
| A(9)= | 0.22755 | 0.01047 |
| A(10)= | 0.33422 | 1.09935 |
| A(11)= | 0.22015 | -0.36053 |
| A(12)= | -0.67517 | 0.61808 |
| A(13)= | 0.15489 | -0.26555 |
| A(14)= | -0.12943 | -0.13633 |

The roots in Z of the above polynomial were found by the above program to be

| | REAL | IMAG | \|LOGMAGNITUDE\| | |
|---|---|---|---|---|
| ROOT(1) | 0.62535 | -0.07880 | 0.46157 | |
| ROOT(2) | 0.35251 | 0.94711 | 0.01053 | DELETE |
| ROOT(3) | -0.21019 | 0.26791 | 1.07728 | |
| ROOT(4) | 0.47505 | 1.68848 | 0.56192 | |
| ROOT(5) | -0.57159 | 0.79850 | 0.01816 | DELETE |
| ROOT(6) | -1.07257 | 0.84329 | 0.31070 | |
| ROOT(7) | -1.76889 | 0.60238 | 0.62521 | |
| ROOT(8) | 0.34753 | -0.64730 | 0.30831 | |
| ROOT(9) | -1.10402 | -0.95889 | 0.38001 | |
| ROOT(10) | -0.34591 | -1.29541 | 0.29327 | DELETE |
| ROOT(11) | 0.22513 | -1.10904 | 0.12369 | DELETE |
| ROOT(12) | 1.95558 | -0.38249 | 0.68946 | |
| ROOT(13) | 2.43781 | -0.97210 | 0.96488 | |

The four roots of magnitude closest to unity were determined by comparing the values of ABS(REAL(CLOG(ROOT(I)))). The complex logarithm function CLOG returns a real part equal to the logmagnitude. The roots with the smallest absolute value of this logmagnitude are ROOT(5),ROOT(2),ROOT(11) and ROOT(10) and were deleted to produce the flattened curve of the reduced determinant.

The adjoint matrix polynomials before dividing by the reduced determinant have the nine frequency responses shown in figure 8A.

When the reduced determinant is used as the denominator and the adjoint matrix polynomials are used on the numerator to obtain the nine-filter transmit

preconditioning, the nine frequency responses of figure 8B are obtained. When combined with the actual propagation paths, these responses contrive jointly to ensure that each mobile receiver receives only its own intended signal through an effective channel comprised of the product of the four deleted determinant factors. Due to selecting the deleted roots to be those closest to the unit circle, this effective channel comprises four rays that are exactly symbol spaced and contain the maximum possible energy.

The nine frequency responses of figure 8 can also be combined in threes by adding their power responses to determine how much power in total is being used to transmit to each mobile receiver, as shown in figure 8C. The integral of the power spectral curves yields the total power used for transmitting the intended signals to each mobile. These powers can be compared to the powers which would have been necessary to communicate the same total signal power one mobile at a time from the best base station, with the same channel state to obtain gain or loss for the macrodiversity system. In the above case, the results were:-

| | |
|---|---|
| Mobile 1: | +4.46dB more power needed (a loss) |
| Mobile 2: | +0.98dB more power needed (a loss) |
| Mobile 3: | +2.50dB more power needed (a loss) |

The net loss is less than appears, because it is the mobiles that are relatively nearer their bases that receive the significantly increased power while those further away receive smaller power increases.

After carrying out many trials using random propagation channels each comprising the exemplary three, symbol-spaced, delayed multipath rays, a histogram was obtained of the total transmitter power used compared to the transmitter power which would have been used for each base station talking to one mobile station independently, ignoring mutual interference. Figure 9 shows the histograms of gain/loss for this 3x3 system, the different curves corresponding respectively to the deletion of 1,2,3,4 or 5 poles from the determinant.
Figure 9A shows the gain/loss histogram for the case where the number of poles deleted was selected in each trial to provided maximum communications efficiency. The mean loss in the 3x3 case is about 1.5dB in this case, for which the benefit of a tripling in capacity is obtained. It will be shown below that overprovision of base station transmitters in relation to mobile receivers can turn this net power loss into a net power gain.

The logmagnitude of a root of the determinant polynomial indicates the rate of decay of the impulse response due to the associated root, in Nepers per symbol period. It could be decided that it was unnecessary to delete roots representing greater than 3dB (0.35 Nepers) decay per symbol period, for example. If only three roots had logmagnitudes less than 0.35, then only those need be deleted. Therefore an implementation of the invention can comprise deleting only those roots with absolute logmagnitude less than a threshold, and thus not always the maximum number that the receiver equalizer can handle, with the intention of approximating the performance of figure 9A for deleting the optimum number of roots.

The key to coherent macrodiversity as described above is knowledge of downlink CSI at the transmitting network. This may be rendered possible under the following circumstances:

(i) Uplink and downlink use the same channel frequency alternately in quick succession, a so-called time-duplex or ping-pong system. Then the transmitter may assume that the downlink channels are the same as it measures on the uplink when it decodes the signals received back from the receivers.

(ii) The receiver measures downlink channel-related information, encodes it and transmits it back to the transmitting network with small turnaround delay. For example, the UMPTS Wideband CDMA system (W-CDMA) has the ability to serve up to 200 voice users per frequency channel per cell, or a proportionally lower number of high bitrate users such as mobile web-browsers. Therefore, for mobile web-browsers desirous of receiving a high instantaneous datarate, it is acceptable to use the whole capacity of a voice channel or more on the uplink to feedback channel state information.

(iii) In a mobile satellite communications system where the relative coupling from transmit antenna elements to receivers is almost static depending only on receiver position.

(iv) In a wireless-in-the-local-loop system for transmitting internet or voice services wirelessly to the home, and the receive antenna is fixed.

(v) In mobile systems where the mobile terminal is likely to be stationary when high bitrate services are invoked.

The above systems all provide feedback of channel state information, but the case of fast-moving receivers is the most challenging as the CSI changes rapidly, and low-delay, high-rate feedback of CSI is required. Some solutions for rapid feedback of changing CSI will now be described.

Let [C'] denote the current CSI assumed by the transmitter, which is in error from the correct CSI [C] by an error matrix [E] so that

$$[C'] = [C] + [E], \quad \text{or conversely } [C] = [C'] - [E]$$

The transmitter transmits $[C']^{-1}PjSj$ where Pj is the effective net channel for signal Sj. Pj is the factor by which prefilters 10A, 10B or 10C of figure 2 differ from the true determinant polynomial.

Receiver(i) receives

$$Ri = [C]_{ik}.[C']^{-1}{}_{kj}PjSj$$

where summation over the common index k is implied

$$= [C'-E]_{ik}.[C']^{-1}{}_{kj}PjSj$$
$$= PiSi - [E]_{ik}.[C']^{-1}{}_{kj}PjSj,$$

since $[C']_{ik}.[C']^{-1}{}_{kj} = d^{ij}{}^{1}$ if i=j else 0.

Thus if receiver(i) correlates its received signal with known symbols embedded in the transmission Sj to receiver(j), the error polynomial term $[E]_{ik}.[C']_{kj}Pj$ summed over index k will be obtained.

If all mobiles do this for all j including their own and return the results to the transmitter, since the transmitter knows the Sj it transmitted, the prefilters Pj it used, and the assumed CSI C'ij, it can compute Eij and hence correct C'ij towards the correct or changing Cij, thereby tracking changes in CSI. Thus the receivers can return measurements of uncancelled interference from symbols intended for other receivers, and delayed versions of the same. From these interference correlations, the transmitter deduces how its CSI must have been in error, and corrects it. Specifically, receiver 1 reports the polynomials determined by correlation with shifts of respective known symbol

patterns as follows:-

$$X11(z) = P1 - \text{sigma}\{E_{1k}C^{-1}_{k1}\}P1$$
$$X12(z) = - \text{sigma}\{E_{1k}C^{-1}_{k2}\}P2$$
$$\cdot \quad \cdot \quad \cdots\cdots\cdots$$
$$X1N(z) = - \text{sigma}\{E_{1k}C^{-1}_{kN}\}PN$$

This is a set of N equations for the N unknown polynomials E11,E12,E13....E1N

Likewise, receiver 2 reports

$$X21(z) = - \text{sigma}\{E_{2k}C^{-1}_{k1}\}P1$$
$$X22(z) = P2 - \text{sigma}\{E_{2k}C^{-1}_{k2}\}P2$$
$$\cdot \quad \cdot \quad \cdots\cdots\cdots$$
$$X2N(z) = - \text{sigma}\{E_{2k}C^{-1}_{kN}\}PN$$

and this is a set of N equations for the N unknown polynomials E21,E22,E23...E2N,

and receiver N reports

$$XN1(z) = - \text{sigma}\{E_{Nk}C^{-1}_{k1}\}P1$$
$$XN2(z) = - \text{sigma}\{E_{Nk}C^{-1}_{k2}\}P2$$
$$\cdot \quad \cdot \quad \cdots\cdots\cdots$$
$$XNN(z) = PN - \text{sigma}\{E_{Nk}C^{-1}_{kN}\}PN$$

which is a set of equations for EN1,EN2....ENN.

The solution of each of such sets of equations for one row of [E] is

$$[C][P_{-1}].X \quad \ldots\ldots\ldots\ldots\ldots\ldots(1)$$

where $[P_{-1}]$ is a diagonal matrix of the reciprocals of the prefilters. If the reported measurements X were exact, the X polynomials would contain P as a factor which would cancel. The remaining factors would give a solution for E that was entirely FIR, i.e. no denominator polynomials, as required. However, due to noise, the reported X polynomials will not have this exact property. A solution is find the pure FIR solution of order L for E that best matches the frequency responses given by equation (1). For example, denominator roots from P can be paired with the closest numerator roots from C or X for annihilation until only numerator roots remain. These then yield the "best" pure FIR solution for E.

Figure 14 shows another method for providing channel state feedback from mobiles to the network. A network comprising network processing (200) and antenna sites (201,202) forms signals for transmitting to the mobile stations using the two sites. Mobile 100 comprises a transmit/receive antenna 101 coupled via a duplexer 102 to receive circuits 103 which filter, amplify and convert the composite received signal to signal samples, preferably in digital form, i.e. using an AtoD convertor. The signal samples from the receiver are then added in 105 to a pilot code and then fed to transmitter circuits 104. The transmitter circuits convert the signal samples to an continuous signal using a DtoA convertor for digital samples, and the continuous signal is upcoverted to a transmot frequency, amplified to transmit power level and transmitted through duplexor 102 connected to antenna 101.

The network sites receive the transmitted loop-back signal from various mobile transmitters. The loopback signals from different mobiles may be

separated by interference rejection combining of the signals from the different sites in processor 200. Also, processor 200 computes correlations between the received signals and the pilot code inserted by a mobile to determine the uplink channel. Correlations are also computed between the network received signal from a mobile and the signals the network transmitted from each of its sites in order to determine the total loopback channel, which is the product of the downlink and uplink channels. The uplink channel is then divided out to reveal the downlink channel. If necessary, the network sites can also each add a different, low-level pilot code to theor transmissions which would be chosen to assist in this loopback channel determination. Using this method, the mobiles are relieved of the complexity of performing channel determination.

The mathematics given above include the derivation of the optimum transmit signals when the number of transmitters exceeds the number of receivers. With such overdimensioning, more degrees of freedom exist to find a way of overcoming fading on any particular path, and/or of cancelling interference at each receiver. This should translate to fewer pathological determinant roots close to the unit circle, and therefore to greater transmit power efficiency. To obtain the benefits of overdimensioning it is only necessary for the number of independently fading paths of the same delay to be greater than the number of receivers. For meeting this criterion, mean delay from an additional base station can always be made the same as the delay from existing base stations artificially, which occurs automatically when the above-described mathematical operations are performed. However, this criterion can also be met by providing, at one or more base stations, transmitters for both polarizations, for example Right Hand Circular and Left Hand Circular. Field measurements confirm that the propagation channel for a transmitted wave of one polarization reaches a receiver over a landmobile propagation path shows little correlation to the propagation channel of the other, orthogonally polarized wave. Thus if each base station site is equipped to transmit power on both polarizations, the base stations each count as two independent base stations as regards the mathematical formulation, even though the RHC and LHC transmitters are co-located.
Figure 10 shows a gain/loss histogram computed from many trials of a 3-base/3-mobile system wherein each base station is equipped to transmit both polarizations. The channel propagation matrix is then a 3x6 matrix of polynomials, and the transmit signal generator computes six waveforms to be transmitted from respective base stations and with respective polarizations, based on the three information signals to be transmitted to respective mobiles. The different curves again correspond to the deletion of 1,2,3,4 or 5 of the "worst" determinant poles, and now show a net gain in total power efficiency of 4-4.5dB in the mean. That means the sum of the power transmitted at all three sites and in both polarizations can be significantly less than needed in prior art systems not using the invention at the same time as allowing the same channel to be reused three times at each location. Moreover, the effect of using both polarizations to thereby overdimension the number of transmitters relative to the number of receivers is to virtually eliminate determinant poles close to the unit circle, so no determinant poles need be deleted. If no determinant poles are deleted, then the system provides a constant net propagation channel for information signals from the network to the mobile receivers, which is free of fading and multipath distortion.

Strategies for slating determinant poles for deletion other than selecting the poles nearest to the unit circle can be used. If for example the adjoint matrix polynomials in the same column all have a factor equal to or close to a determinant factor, then it is unnecessary to delete that determinant factor as it cancels with the same factor in the numerator. Likewise, when the numerators contain a zero close to an offending pole, they reduce the undesirable effect of the offending pole. A systematic way to detect and exploit such occurences to select the poles to delete for best effect is desired. This systematic method

is provided using Cauchy's Residue theorem. Cauchy's residue theorem provides a much faster method to determine the total power transmitted when a "white" signal input is filtered by one or more numerator (FIR) z-polynomials and a denominator (IIR) polynomial to obtain the transmit signals.
The method is outlined in an Appendix, and may be extended by a mathematician using the same principles to cover all practical cases. In the Appendix, a version of Cauchy's residue theorem is developed for integrating the modulus squared of a complex Numerator frequency function divided by a complex Denominator frequency function. In the case where the Numerator is a z-polynomial at least one order less than the Denominator z-polynomial, the result is a sum of terms, each corresponding to the contribution from a respective denominator pole. If the numerator is small at that pole, then the contribution (the residue at the pole) to the total power integral will be low. Therefore the systematic method desired above is to select for deletion poles that have the largest real part to their residues, inlcuding the numerator term when computing the residues.
It can be argued that the selection process should first select the pole having the largest real part to its residue for deletion, and then re-evaluate the residues of the remaining poles having deleted the pole. This process should continue until the desired number of poles have been deleted. The remaining denominator order must be at least one higher than the numerator order.
If the process shall be continued to denominators of lower order, then an extra residue appears due extra powers of of z on the denominator. The extra residue can be determined by making a partial fraction expansion to extract a term having that power of z on the denominator, the other term now having a numerator of lower order than the denominator. The extra residue is then equal to the coefficient of the highest power of z in the numerator above the extracted term. The extra term is needed to determine the power after deleting a pole causing the numerator order to equal or exceed the denominator order.

In the application to this invention, the numerator can comprise the sum of the moduli of all the adjoint matrix polynomials down a column, where the modulus of a z-polynomial means the product of itself with its time-reversed conjugate, as further elaborated in the Appendix. The residue at a pole therefore comprises a part from each adjoint matrix numerator polynomial, so only if all adjoint matrix polynomials in the same column are small at the denominator pole with that pole have a small residue, and thus avoid being slated for deletion.

When practising the invention, it is desirable to group mobiles using the same channel according to their locations repative to a group of three neighboring base stations. Figure 6 illustrates the difference between desirable and less desirable groupings. The desirable groupings in figure 6 are obtained by forming group 1 as comprising the three mobiles which nearest to their respective base stations. This can be considered as producing a propagation loss matrix with the least loss along the diagonal. Then group 2 comprises the three mobiles with the second least loss to a respective one of the three base stations; group 3 comprises the three mobiles with the third lowest loss to their respective base stations, and so on. The least desirable groupings would comprise grouping together the three stations nearest to a single base station, for example. The reason for preferring the former grouping method is that it obtains the highest possible wanted-to-unwanted signal ratio even without employing the invention, such that the additional signals transmitted to cancel unwanted cochannel interference when practising the invention are a minimum. With the least desirable grouping, three mobiles receive cochannel signals from the same, nearest base station, and the other, more distant base stations must transmit a lot of interference-cancelling signal power due to their greater distance.

It is always desirable to simplify mobile receivers such as mobile cellular phones due to their large numbers, and place complexity instead in the network stations, which are much less numerous. Thus a simplified method by which the receivers can feedback downlink channel information to the transmitting network would be useful. For example, the signal received at each receiver could be simply turned around and retransmitted with minimum delay back to the network, as shown already in figure 14.

If all receivers do so on the same channel, the network must separate them by uplink beamforming/interference cancellation, which implies knowledge of uplink CSI. Uplink CSI is also needed to divide out the effect of the uplink channel polynomials on the retransmitted signal so that it reflects only the effect of the downlink channel. In a CDMA system, the mobile stations can retransmit the signal received on the downlink with the addition of an uncorrelated pilot code sequence that the network stations can use to derive uplink CSI. In a non CDMA system that would not tolerate an overlapping pilot sequence, the feedback signal can instead by periodically interrupted at known times to insert pilot symbols that the netwotk stations can use to derive uplink CSI. Thereby the onus for analysing what the receivers have received is placed back on the network. The network has the great advantage of knowing every symbol that was transmitted to every receiver and what prefilters were used for all the signals. The network can therefore perform correlations using the entire symbol sequence transmitted to each receiver, including data symbols and not just known pilot symbols. Thus a modification to figure 14 can comprise interrupting the loopback signal to insert pilot symbols, replacing the additive combination of pilot and loopback signals formed by adder 105. In general, any suitable combination of the loopback signals with mobile-specific pilot symbols or mobile-discriminating information can be used.

Many variations of the above principle of "mirror reflection" of the received signals back to the network can be devised. For example in a CDMA system, the received signal can be despread using the codes of each receiver to obtain despread symbols, then the despread symbols can be respread using corresponding uplink codes and added. The multi-code uplink signal is then mirrored to the network. Interference correlations (the X polynomials in the above notation) can also be digitally coded of course, and transmitted as a data stream protected by error correction coding. For high symbol rates giving long channel polynomials (large L) or for large N (e.g. greater than 3) the amount of digital information to be transmitted may exceed the uplink capacity available, presumed to be for example the capacity of one voice channel, or about 4 to 12 kilobits per second. The information could be selectively reduced by including in the reports only the X polynomial or polynomials having the greatest coefficient magnitudes; only polynomial coefficients that had changed by more than a threshold amount from a predicted value, or othe rmeans of down-selecting. Reporting only the coefficient with the greatest magnitude will cause the network to correct the transmitted signals to reduce only that largest interference component, which however if repeated successively will sequentially reduce the interference in order of strongest components first.

Another implementation of the invention with reduced CSI reporting requirements comprises the mobile stations reporting only the quality of the signal they are instantaneously receiving from the multi-antenna transmitting system. The transmitters transmit signals using a fixed or systematically varying phasing. The network examines the signal strength reports from the receiving stations and selects to transmit data intended for the receiver that is instantaneously reporting the highest quality increase over its mean received quality. Due to channel variations and/or the above-mentioned systematic variations in macrodiversity phasing, it would be expected that each receiver received the downlink capacity for an equal fraction of the time.
Thus, according to this embodiment, a group of for example three network

transmitters transmit respective modulated signals derived from the same information symbol stream. The modulated signals are derived from the information symbol stream by varing the relative delay of the information symbol stream for at least two out of the three modulation signals with respect to the third, such that, taking account of different propagation delay to the receiver for the dominant multipath ray, the modulation is received in synchronism at some receiver. Alternatively, the phase of the modulated radio frequency signal can be altered separately for at least two of the three network stations such that, taking account downlink channel phase for the dominant ray, the dominant rays are received in phase at some receiver. Preferably both phase and delay are altered for two out of three transmissions so that for some receiver, the dominant rays are received in synchronism and in phase.

The receivers all demodulate the received signal and determine the quality of the demodulated data, which can for example be the total signal power formed from the sum of the squares of the ray amplitudes given by downlink channel estimation. The receivers report the measured signal quality continuously, for example by mirroring a signal related to the measured quality, combined with pilot symbol sequences to provide uplink CSI to the network. The network monitors the mirrored signal quality reports and chooses the data that it is instantaneously transmitting to be the data intended for the receiver intantaneously reporting the best signal quality relative to its mean reported signal quality. The number of receivers reporting signal quality in such implementations is not related to the number of transmitting antennas.

One example of a macrodiversity transmission using the above principles would comprise three network transmitters, each transmitting the same information symbol modulation with no relative delay, but with relative frequency offsets of 0, -20Hz and +20Hz. The frequency offsets assure a varying net channel even if the receivers are static.

Yet another implementation comprises a network of exemplary three neighboring transmitters. An unrelated number of receivers receives the signals from the three transmitters and each receiver reports the transmitter it receives most strongly and the signal strength. The network selects to transmit data to the receiver receiving the strongest signal relative to the mean reported signal strength, using the best transmitter. The need for full, complex downlink CSI is avoided in these implementations because no attempt is made to perform coherent interference cancellation in order to transmit three signals to three receivers at once. The above-described methods are different than the known method of choosing the destination receiver based on its instantaneous reception quality. These known method perforemd selection among receivers already assigned to a single base station, as opposed to the above methods which dynamically reassign the mobiles to receive from different base stations or conversely, dynamically reassign the base station that shall transmit to each mobile.

In dynamically "re-sorting" the assignment of one of three base stations to one of three mobiles, there is a previously unreported gain in communications efficiency which is computed and reported below.

In the prior art, a mobile station indicates its preferred base station and is assigned to that base station for a session. If fading temporarily hinders transmission, that base station is preferably temporarily used to transmit data to an unfaded mobile. The prior art does not attempt to do the above while using the same channel three times in an area served almost alike by three base stations. If this were attempted, the mean wanted-to-unwanted signal levels at each mobile (C/I's) may be computed from the mean distance loss using a 4th power propagation law, and averaged over a period long compared to the Rayleigh fading. On the other hand, when practising the current invention, the base stations reassign themselves to the three mobile stations dynamically

using the feedback or loopback information in such a way as to minimize the total power transmitted for maintaining a unit received signal level at each mobile. The average C/I's are then computed for the invention, and the following comparison obtained:

| | Mobile 1 | Mobile 2 | Mobile 3 |
|---|---|---|---|
| C/I without adaptive allocation | 4.47dB | 4.93dB | 5.12dB |
| C/I with adaptive allocation | 4.81dB | 5.31dB | 5.14dB |
| Gain for adaptive allocation | 0.34dB | 0.38dB | 0.02dB |

The above gains were obtained with the relative base and mobile positions shown in figure 1. The mobiles received data from their nearest bases in the above tests 61.3% of the time, and fading caused re-allocation 39.7% of the time.

It is interesting to place the mobile stations in greater proximity to eachother where the C/I would be expected to be very low with conventional fixed allocation. With three mobiles nearly equidistant from the three base stations and within 100 meters of eachother, the following results were obtained:

| | Mobile 1 | Mobile 2 | Mobile 3 |
|---|---|---|---|
| C/I without adaptive allocation | -1.65dB | -1.61dB | -1.62dB |
| C/I with adaptive allocation | 2.26dB | 2.89dB | 2.08dB |
| Gain for adaptive allocation | 3.91dB | 3.50dB | 3.70dB |

With three almost colocated mobiles, the allocation of base stations to mobiles stations exhibited almost equal probability of each of the six possible assignment arrangements. The arrangement that occurred most frequently during adaptive assigment was used as the baseline for the non-adaptive case.

Thus there is a considerable mitigation of interference between almost colocated mobile stations when practising dynamic base station reallocation according to the invention. The invention guarantees a mean C/I of around 2dB minimum as opposed to approximately -2dB without the invention.
This translates to a more than doubling of achievable data rates together with use of the same channel three times over in a given area. This gain may be achieved by reducing the spreading rate of a CDMA system for example, to take advantage of the C/I improvement by providing a higher data rate. Alternatively, the amount of error correction coding used can be reduced when the invention is employed, also increasing the useful information rate while maintaining acceptable error rates. The above advantages are obtained when practising fast, adaptive allocation during a transmission as opposed to the prior art of "slow" adaptive allocations made prior to the start of transmissions. Fast adaptive allocation switches base-mobile allocation in response to Rayleigh fading patterns, while slow adaptive allocation switches base-mobile allocations only in response to mean path-loss changes.

In another implementation differing from the prior art, when a base station selects not to transmit to a faded mobile station, the faded mobile station may receive data instead from a base station to which it is unfaded, the network dynamically determining the base station to transmit the data based on the feedback signals. Alternatively, the network determines a mobile station and a base station in the vicinity that may momentarily use the channel without interference, based on which base-mobile link currently is the least faded relative to an equalizing threshold. The equalizing threshold is chosen for each mobile station to provide approximately equal probability of each mobile station receiving data, and therefore equal mean data rate.

A degenerate case of the M-antenna, N-receiver solution is the use of two antennas, for example two colocated RHC and LHC transmitting antennas, to transmit a signal to a single receiver. When the two antennas are colocated and differ only by polarization, the channel coefficients for the multipath channels are substantially uncorrelated for the two polarizations; however, the delays of the fading, multipath rays are the same. That is, if the channel for RHC is expressible as

$$C^{RHC}(Z) = C_o^{RHC} + C_1^{RHC} Z^{-P1} + C_2^{RHC} Z^{-P2}$$

then the LHC channel is expressible as

$$C^{LHC}(Z) = C_o^{LHC} + C_1^{LHC} Z^{-P1} + C_2^{LHC} Z^{-P2}$$

The power of Z are the same for both polarizations, expressing equality of the ray delays, but the coefficients are different and fade in an uncorrelated fashion.

The optimum way to transmit a wanted signal to the receiver is then given as before by

$$C^{\#}[C.C^{\#}]^{-1}$$

times the wanted signal waveform, where C is now the 2 x 1 matrix of polynomials

$$C = \begin{pmatrix} C^{RHC} \\ \\ C^{LHC} \end{pmatrix}$$

Thus $[C.C^{\#}]^{-1} = 1/(C^{RHC}.C^{\#RHC} + C^{LHC}.C^{\#LHC})$

and there is no adjoint matrix. The implementation then depends only on the location of the poles of this denominator/determinant polynomial. It may be appreciated that, when P1 and P2 are large integers, this determinant has a prodigious number of roots even though it has a small number of coefficients.

The above channels, which have three significant multipath rays of relative delays 0, P1 and P2 symbol periods, are typical of a CDMA channel. A known CDMA receiver would conventionally employ a form of multipath equalizer known as a RAKE receiver, which correlates the received signal with a locally generated despreading code shifted in time by 0, P1 and P2 symbols. The correlation results are then weighted and added. The number of delays used and correlations combined is called the number of "RAKE taps". The RAKE taps in the present example are placed at delays of 0, P1 and P2 symbols or chips. The RAKE receiver is not limited so much by the amounts P1, P2 of delay, but in the total number of taps. Thus, if a three-tap RAKE receiver were provided for use without the inventive coherent macrodiversity transmit scheme, it would be desirable that no more taps should be needed with the invention. However, omitting the whole of the above determinant polynomial would result in a net transmission channel having seven delays and needing in principle a 7-tap RAKE receiver with tap delays of

-P2, -P1, -(P2-P1), 0, (P2-P1), P1 and P2 chip periods.

If the already-described pole-deletion procedure is used to improve the conditioning of the denominator for implementation, only two roots may be deleted while preserving only a 3-tap channel for the RAKE receiver to deal with. Moreover, the three tap channel would always comprise delays of exactly

0, 1 and 2 chips with this method. Thus the three RAKE taps would be adjacent instead of at 0, P1 and P2 symbols delay. However, the RAKE receiver is limited only in the number of taps available to combine, and not their spacings. Therefore any three-coefficient numerator polynomial of the form

$$Ao + A1.Z^{Q1} + A2.Z^{Q2}$$

could be used as an FIR prefilter, and would preferably share roots with the denominator that it is desired to cancel. Alterntively, a polynomial in which most of the energy was contained in three taps could be used.

A denominator polynomial of the form $C_1(z)C_1{}^{\#}(z) + C_2(z)C_2{}^{\#}(z) + \ldots\ldots$ has the property that half of its roots are the conjugate reciprocals of the other half. Those roots having magnitude greater than unity ( and which would therefore require time-reverse IIR processing) may be slated for deletion. The remaining poles form a causal IIR filter which may be implemented entirely by forward-time processing, and the deleted poles give rise to a net FIR channel. This is one way of factorizing the denominator polynomial into the form $C_{equ}(z)C_{equ}{}^{\#}(z)$, which factorization is always possible by selecting one root to form $C_{equ}$ and its conjugate reciprocal to form $C_{equ}{}^{\#}$. The choice of the larger or smaller root for $C_{equ}$ can be made in $2^{N/2}$ different ways, given N/2 pairs of roots, so the number of such factorizations that can be made is large.

Trial computations showed that factorization into a purely causal and a purely anti-causal part gives a channel, although not limited to the same number of taps as the original channels C1 and C2, nevertheless contains most of its energy in relatively few coefficients, allowing a RAKE receiver with a small number of taps to be used. When fewer taps are used than the number of effective channel rays containing signal energy, a loss of energy is entailed. Figure 12 plots the PDF of this energy loss for using 1 to 6 taps to receive the signal through the above-mentioned anti-causal factor. The RHC and LHC channels were both of the form

$$Co + C1.z^5 + C2.z^{13},$$

and thus the determinant is of order 26. The causal and anti-causal factors are then both of order 13, but do not have non-zero z-powers of only 5 and 13. Nevertheless, 80% of the energy is most often contained in the tap delay 14, 93% is contained in tap delay 14 and tap delay 9 and 98% is most often contained in tap delays 14, 9 and 1. Thus it appears that a RAKE receiver with no greater a number of taps than would have been used for the conventional channel will be more than adequate for use with the invention. Figure 13 plots a histogram of the net gain in communications efficiency for this method, when the number of RAKE taps is fixed at three, the same as the number of rays in the multipath channels. The gain of 2dB equates to a 60% increase in communications capacity in terms of number of users or datarate when practising the invention.

Figure 15 shows another application of the invention to a CDMA system.

The signal to each mobile is spread-spectrum coded using a unique spreading code, and the set of spreading codes may be mutually orthogonal, at least when time-aligned. When not time-aligned, such codes will show some degree of correlation or mutual interference.

The conventional CDMA system comprises inputting symbol streams $S_1 \ldots . S_N$

to spread spectrum encoder 300, where $S_i$ is the coded information stream intended for mobile(i). Stream $S_i$ is multiplied with spreading code $C_i$ and transmitted to mobile(i) through downlink channel polynomial $P_i(z)$.

Mobile(i) correlates the received signal with a first shift of code $C_i$ and obtains $Po(i).S_i$ where $Po(i)$ is the first coefficient of $P_i(z)$ and likewise obtains $P1(i).S_i$ for other code shifts.
These are then RAKE combined into a single sample stream whose dominant component is then, hopefully, $Qo_{ii}.S_i$
where $Qo_{ii} = |Po|^2+|P1|^2 \,..|P_{L-1}|^2$
and L is the length of the downlink channel to mobile(i).

However, the combined RAKE output signal will contain other terms. There will be some Intersymbol Interference from the previous and the next symbol of stream $S_i$ due to correlation of mobile(i) code shifted with itself. If the channel time dispersion is less than one information symbol in duration however, only one previous and one future symbol will contribute to the RAKE output. This can be accounted for by coefficients $Q_{-1}$ and $Q_{+1}$ of a 3-coefficient z-polynomial $Q_{ii}(z)$ and the combined RAKE output can be written

$$Q_{ii}(z).S_i(z)$$

In addition, there are correlations between code(i) of mobile(i) and code(j) of mobile(j) shifted, which give terms $Q_{ij}(z).S_j(z)$.

The order of $Q_{ij}$ may, as already seen in the above simplified example, be extended when signal(j) is received from a base station lying at a greater or lesser distance from mobile(i), but not when all signals are received from the same base station. Thus all $Q_{ij}$ in this example have three coefficients. This matrix of z-polynomials described the coupling from other mobile's intended information symbols to a given mobile's RAKE output, and is not precompensated at the transmitter in the prior art. Using the invention, the coupling matrix of z-polynomials may be inverted by forming the adjoint matrix and the determinant polynomial.
If the determinant polynomial has no unpleasant roots, then using the inverted matrix as precombiner 301 cancels all interference at the receiver output of each mobile from both other mobile's symbols and from its own previous and future symbols. If unpleasant roots of the determinant polynomial have to be deleted, and the remaining, reduced determinant used as the IIR part of precombiner 301, then all interference from other mobiles' signal will still be cancelled, but there will be a net multipath channel created from the product of the omitted determinant factors through which each mobile will receive its intended symbol stream. This can be completely compensated at the mobile receivers by following the RAKE combiner output with a 3-tap, Viterbi Maximum Likelihood Sequence Estimator or other known form of symbol-spaced equalizer running only at the information symbol rate and not at the spread spectrum chip rate. Thus figure 15 contemplates the addition of an information-rate matrix precombiner to combine information symbol streams s1,s2...sn to form information-rate sample streams $S1, S2....S_N$ which are input to the conventional spread spectrum transmitting device 300. The precombiner 301 has the same structure as already shown in figure 2. The polynomials used in precombiner 301 are computed using knowledge both of the downlink channels $P_i(z)$ and of the cross-correlations between and and the autocorrelations of the different spreading codes used in coder 300,

and using the assumption that the mobile receivers employ conventional RAKE combining with weighting coefficients adapted to the chip-level channel impulse responses.

The coupling or interference matrix Cij is defined as follows.

figure 16a illustrates how a RAKE output sample depends typically on three adjacent symbols, due to delayed multipath rays. When the sample corresponding to symbol (i) is calculated by the RAKE receiver, it first correlates the received signal with the chip sequence used to spread symbol(i), with the chip sequence aligned in time with ray 1 to obtain a value U1, then with the chip sequence time aligned with ray2, ray3, ray4... to obtain values U2,U3,U4... correspondingly. Correlation means multiplying each received chip sample by the complex conjugate of the the spreading code symbol, and summing over the number of chips used to spread the symbol to obtain each U-value. The U-values are then weighted by the complex conjugates of the separately estimated channel coeffcients C1,C2,C3,C4... for rays 1...4, giving the RAKE receiver output R(i) as

$$R(i) = C1^{\#}U1+C2^{\#}U2+C3^{\#}U3+C4^{\#}U4...$$

However, as may be seen from figure 16a, when correlating with the chip sequence time-aligned with ray 1, there is an overlap with the end of the chip sequence for symbol(i-1) due to delayed rays 2,3 and 4, so that there is a contribution to U1 from Symbol(i-1). This contribution is

$$S(i-1)[C2.p2+C3.p3+C4.p4..]$$

where p2 is the partial correlation between the left-hand end of the chip sequence for symbol i and the overlap with the right-hand end of the chip sequence for symbol(i-1) due to delayed ray2;
p3 is the partial correlation of the chip sequence for symbol i for the slightly greater overlap with the end of the chip sequence for symbol(i-1) due to delayed ray3, and so on.

Likewise, U2 has a contribution from S(i-1) due to delayed rays 3 and 4 of

$$S(i-1)[C3.p2+C4.p3..]$$

and U3 has a contribution from S(i-1) due to delayed ray4 of

$$S(i-1)[C4.p2..]$$

Thus the total contribution to R(i) from S(i-1) is

$$C1^{\#}.[C2.p2+C3.p3+C4.p4..]+ C2^{\#}.[C3.p2+C4.p3..]+ C3^{\#}.[C4.p2..]$$

which is therefore the first coefficient $C11^{-1}$ of the channel polynomial C11 that describes interference between a mobile symbol and the preceding symbol for the same mobile, i.e. ISI at the output of the RAKE receiver. Likewise, $C11^{+1}$ describes ISI from the succeeding symbol S(i+1) and is given by

$$C4^{\#}.[C3.q3+C2.q2+C1.q1..]+ C3^{\#}.[C2.q3+C1.q2..]+ C2^{\#}.[C1.q3..]$$

where now q3 is the partial correlation between the overlap of the right-hand end of the code sequence for symbol(i) with the left-hand end of the code sequence for symbol(i+1).

The middle coefficient $C11^{\circ}$ of polynomial C11 is given by

$C1^{\#}.[C1.w1+C2.w2+C3.w3+C4.w4..]+$
$C2^{\#}.[C1.w2^{*}+C2.w1+C3.w2+C4.w3..]+$
$C3^{\#}.[C1.w3^{*}+C2.w2^{*}+C3.w1+C4.w2]+$
$C4^{\#}.[C1.w4^{*}+C2.w3^{*}+C3.w2^{*}+C4.w1..]$

and the value of w1 is unity, as it is the correlation of a code with itself fully aligned.

The above exemplary expressions are for the case of four delayed rays, but may be extended to any number of rays in an analogous manner. Likewise, if ray delays of greater than one symbol period are encountered, the channel polynomials such as C11 may lengthen to 5 coefficients to incorporate dependence of the RAKE output sample on S(i-2) and S(i+2).

The entire matrix of polynomials Cij is defined in the same manner as above, except that Cij means the dependence of a sample from RAKE receiver(i) (which uses mobile(i)'s code, code(i), for correlation) on the symbols of mobile(j) which are spread with code(j). The values of the partial correlations above are then between overlapping segments of different codes, and in particular, the value of w1 is no longer unity, but the correlation between a symbol-length segment of code(i) and a fully aligned segment of code(j), which is zero when codes (i) and (j) are orthogonal.

If the multipath delay spread is less than an information symbol in duration, each sample stream will depend only on current information symbols, one symbol period delayed symbols and one symbol period advanced symbols, making each Z-polynomial of order three.

Figure 16a is for the case of transmitting from one base station to 'N' mobile stations. The matrix C is then square and of size NxN.

For the case of transmitting from M diversity base stations to N mobile stations, the total number of inputs to the M base station transmitters would be MN, so the corresponding interference or coupling matrix C, which describes the coupling from each transmitter input to each mobile RAKE receiver output, would be of dimension MN x N. Since this is an overdimensioned case, the optimum transmit conditioning is then given by

$$T = C^{\#}[C.C^{\#}]^{-1}S$$

where C is the above-defined matrix of MN x N coupling polynomials, S is a vector of information symbol sequences intended for the receivers, and T is a vector of as yet unspread sample sequences for transmission. Each of the unspread sample sequences comprises a sequence of samples at the information symbol rate, which samples however are not necessarily of the same values as information symbols, but are multi-valued samples which are additive combinations of many symbols. Each sample, being of symbol-period duration, is then multiplied by the spreading code segment for that symbol corresponding to the transmitter input to which it is applied. The results for all inputs of the same transmitter are added and transmitted as a composite spread-spectrum (CDMA) signal. When all M transmitters collaborate in this way, the RAKE output of receiver (i) contains the intended information symbol stream S(i) only, with neither ISI nor interference from symbol streams intended for other receivers. If instead of $[C.C^{\#}]^{-1}$, $[C.C^{\#}]^{adj}$ is used instead, the RAKE output will be free of interference from non-intended symbol

streams, but will contain ISI. The ISI delay spread will depend on whether all of the determinant of $C.C^{\#}$ is omitted from the inverse, or whether only ill-conditioned poles are omitted. The advantage of the overdimensioned case is that all the poles of the determinant lie further from the unit circle, and therefore are more well-conditioned.

The invention may also be applied to the CDMA uplink, i.e. receiving N mobile signals using one or more base station sites and coherently cancelling interference from one mobile transmission to another, using only symbol-rate processing of the base stations' RAKE outputs.

Such coherent uplink macrodiversity reception relies on the determination of the uplink channels from mobiles to base stations which is more straightforward, and does not rely on fedback information on the downlink channels. Uplink channel estimation may be carried out for each mobile signal at each base station using known pilot symbols included periodically by the mobiles in their transmissions, or using pilot codes radiated by the the mobiles superposed on their information-modulated transmissions, or may be estimated from information modulated and spread spectrum coded transmissions by feeding back decoded symbols (decision feedback channel estimation) or any combination of these methods.

Each base station implements a RAKE receiver for each mobile signal matched to the estimated channel for that mobile signal to produce a complex sample stream of one RAKE combined sample per information symbol for each mobile at each base station. These samples are transmitted coherently to a central receive signal processing center along with channel information. Using the uplink channel estimates and the known spreading codes, the signal processing center can model each complex sample stream's dependence on the information symbols of a wanted mobile signal as well as the interference added to the sample stream from the other mobile signals, thus forming a matrix C of Z-polynomials. This is calculated by computing correlations between the spreading codes and combining the correlations with the channel estimates.

For receiving N mobile signals at each of three base stations, the matrix C is of dimensions 3N x N of third-order Z-polynomials. Interference-rejection combining the sample streams can be achieved by matrix multiplying a vector of the RAKE output sample streams by $(C^{\#}C)^{-1}C^{\#}$; however, as already discussed at length above, the inverse matrix comprised of the adjoint divided by the determinant may have determinant poles undesirably close to the unit circle. The function of the three terms in

$$(1/\det(C^{\#}C)).(C^{\#}C)^{adj}.C^{\#}$$

can be equated to

-Matched filtering or optimum diversity combining (the rightmost term $C^{\#}$)

-Interference cancellation between DIFFERENT mobile transmissions $(C^{\#}C)^{adj}$

-Cancellation of intersymbol interference between symbols of the same transmissio ( the leftmost term $1/\det(C^{\#}C)$ )

It is the latter which is problematic. However, instead of cancelling or removing ISI from the sample stream by dividing by the determinant to form a channel-inverse equalizer, the Viterbi Maximum Likelihood Sequence Estimator can be used instead to decode symbols subject to ISI between adjacent symbols.

The Viterbi MLSE hypothesizes symbol sequences and applies the ISI to the symbol sequence to generate an expected sample stream. The sumsquare errors between expected and actual signal samples is then a measure of the correctness of the hypothesis and the Viterbi algorithm determines the sequence that minimizes the sumsquare errors. Providing the noise on successive signal samples is uncorrelated, that is "white", this is also the sequence most likely to be correct. It is known in the prior art that, if the noise is not uncorrelated but "colored", a noise-whitening filter can be used ahead of the Viterbi equalizer. In this application, the color of the noise can be determined by assuming that the noise received at each base atation is white, but then is modified by combination with the Z-polynomial matrices which are known. The noise whitening filter and division by the determinant are thus rolled together into a diagonal matrix of filters $n_{ii}(z)$, the purpose of which are to whiten the noise for each sample stream ahead of the associated Viterbi Equalizer.

The arrangement described above is illustrated in figure 16. A first base station antenna site receives CDMA signals from N mobiles and implements a known RAKE receiver to produce N corresponding output sample streams. Typically, a mobile terminal would be located somewhere within a triangle formed from the three nearest base stations, so figure 16 illustrates also that a second and third base station may receive the same N mobile terminal signals and also implement RAKE receivers for those signals. In the case that more than one base station participates in receiving at least one mobile signal, the participating base stations preferably pool their RAKE-combined sample streams at central processor (300), which is also provided with multipath channel estimates formed within the RAKE receivers. The central processor (300) implements the above described mathematics of z-polynomial matrix manipulations including all or any of forming an adjoint matrix of Z-polynomials; forming a determinant polynomial; deletion of high-Q determinant roots; determination of noise-whitening filters (301) and implementation of per-terminal equalizers (302). Although figure 16 illustrates three base stations producing RAKE-combined outputs for all N terminals, the invention is operable when base stations produce RAKE combined outputs for different numbers of terminals, the resulting matrix then having null-polyomials in the positions where a base station does not produce a RAKE-combined output for any mobile terminal. A base station may also produce a RAKE combined output for a terminal signal that is not within the group of N decoded by central processor (300), but which is sent to a different central processor associated with a different triangle of base stations. Advantageously, each base station may also receive mobile signals using two antenna polarizations and may RAKE-combine the two polarizations locally to produce still N RAKE-combined output sample streams or preferably may RAKE combine the two polarizations separately to produce 2N RAKE combined sample streams such that central processor (300) now has access to a matrix of size augmented to 6N x N polynomials, thus raising the order of diversity reception from 3 to 6. The higher is the order of diversity reception, the lower is the likelihood of encountering determinant roots of excessive Q factor and the lower is then the likelihood of having to delete many determinant roots. This reduces the complexity of the equalizers (302), alternatively allows the use of shorter symbol periods thereby achieving higher datarates.

The inventive CDMA interference cancellation system disclosed above is characterized by the advantage of being able to employ only information-symbol rate processing of the noisy RAKE-combined signals rather than the much higher CDMA chip-rate processing of non-RAKE combined signals in order to achieve diversity reception and interference cancellation at the same time.

A person skilled in the art will be able to employ the teachings above to make many variations of the invention without departing from the spirit and scope of the invention as described in the attached claims.

# 4. CLAIMS

SEE MY EXEMPLARY CLAIMS AT THE BEGINNING

OTHERS AT PATENT ATTORNEY'S SUGGESTION

EXHIBIT

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ERICSSON

March 20, 2001

**VIA FEDERAL EXPRESS**
David E. Bennett, Esq.
Coats & Bennett
1400 Crescent Green
Suite 300
Cary, NC 27511

RE: **REQUEST TO PREPARE AND FILE PATENT APPLICATION**

| | |
|---|---|
| Ericsson Docket No.: | P12622-US1 (BMOT) |
| Title: | **INTERFERENCE CANCELLING CDMA RECEIVING SYSTEM** |
| Inventor: | Paul Dent |
| Responsible Paralegal: | Maura DuBois |
| Responsible Attorney: | Stephen Calogero |

Dear David:

**THIS CASE IS TO BE ADMINSTERED ACCORDING TO ERICSSON'S FIXED FEE BILLING PROCEDURE, AS ESTABLISHED BY ERICSSON LETTER OF DECEMBER 16, 1999.**

Please prepare and file a patent application in the U.S. Patent and Trademark Office on the above-referenced invention. A copy of the invention disclosure and inventor's questionnaire is enclosed. Any potential statutory bar dates should be identified in these documents; however, any potential bars should also be verified with the inventor.

**All rights in this application are to be assigned to Ericsson Inc., a Delaware Corporation, having registered offices in Research Triangle Park, North Carolina.**

If you wish to discuss any matter regarding this application, please do not hesitate to contact the responsible paralegal at (919) 472-1901. Please contact the inventor for any technical assistance in preparing the application. A first draft of the application should be submitted to the inventor within three (3) months of the date of this letter. Please advise the responsible paralegal if there is any problem with this schedule.

Very truly yours,

Maura E. DuBois

Maura E. DuBois
Paralegal – Intellectual Property

Enclosures
cc: Nancy Ferguson
Paul Dent

2001-075

**Ericsson Inc.**
8001 Development Drive
P.O. Box 13969
Research Triangle Park, NC 27709

Coats & Bennett P.L.L.C.

EXHIBIT

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1
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Patents, Trademarks, Copyrights, Trade Secrets, Licensing, and Related Litigation

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June 22, 2001

Mr. Paul W. Dent
**Ericsson Inc.**
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**RE: U.S. Patent Application**
**Ericsson Ref. No.: P12622-US1 (BMOT)**
**C&B Ref. No.: 4015.980**
**Inventor(s): Paul W. Dent**
**INTERFERENCE CANCELATION IN A CDMA RECEIVING SYSTEM**

Dear Mr. Dent:

Enclosed please find the first draft of the above referenced application. Please review and fax me any changes or comments you may have.

Please do not hesitate to call if you have any questions.

Sincerely,

David E. Bennett

David E. Bennett

cc: Maura DuBois

DEB/rwg
Enclosure